

# A 1- to 10-GHz Downconverter for High-Resolution Microwave Survey

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*A downconverter was designed, built, and tested for the High Resolution Microwave Survey project. The input frequency range is 1 to 10 GHz with instantaneous bandwidth of 350 MHz and dynamic range of 125 dB/Hz. Requirements were derived for the local oscillators and special design techniques were implemented to achieve the high degree of spectral purity required.*

## I. Introduction

The 1- to 10-GHz RF-to-IF downconverter (RID) was designed to perform the RF-to-IF frequency conversion portion of the High Resolution Microwave Survey (HRMS) RF system. The RF system is composed of a series of antenna feeds, hybrids, and low-noise amplifiers (LNAs) that cover the band of 1 to 10 GHz and deliver two polarization channels to the inputs of the RIDs. The RID is duplicated to process two channels simultaneously. The IF output is designed to be transmitted over a high dynamic range fiber-optic link (FOL) to the control room, where it is fed into an IF-to-baseband (IF/BB) frequency converter followed by a real-time 32-million-channel fast Fourier transform (FFT) analyzer [1] and sophisticated signal detection algorithms. This article describes the RID as well as the design process and rationale that led to its realization.

## II. Requirements

The task of the HRMS project is to search for extremely weak narrow-band signals over a wide frequency range in

the presence of very strong interfering signals. The challenge in this design was to trade off the competing requirements of wide bandwidth, high dynamic range, and high spectral purity. The requirements for the RID were derived from the RF system requirements and are listed in Table 1.<sup>1</sup>

## III. Design Approach

To assist in achieving a design that meets these challenging requirements, two computer-based design tools were developed. The first tool is a Fortran program that is highly interactive with the user, performing mixer analysis to discover all the intermodulation products resulting from a particular frequency conversion scheme. The program inputs are the RF and local oscillator (LO) frequency ranges and increments, the IF filter shape, the mixer intermodula-

<sup>1</sup> G. A. Zimmerman, *Search for Extraterrestrial Intelligence Microwave Observing Project Sky Survey Element, Sky Survey Operational System, Functional Requirements Document*, Document No. 1720-4100 (internal document), Jet Propulsion Laboratory, Pasadena, California, June 10, 1992.

tion table, the required suppression, and the highest-order harmonic to be calculated. The program outputs a list of all the frequency spurs that result from this mix with suppression levels that exceed the requirement. Figure 1(a) is an example run showing that no intermodulation products violate 60-dB suppression for the first frequency conversion range. Figure 1(b) is the same frequency conversion with a more stringent suppression requirement. The program lists both the frequency components inside the IF band that are suppressed by the mixer as well as the components outside the IF band that are suppressed further by the bandpass filter.

The second tool is a custom spreadsheet that analyzes receiver performance component by component. The inputs to the spreadsheet are the receiver components in sequence with their associated gain, noise temperature, bandwidth, and compression point. The outputs of the spreadsheet are the contributions of each component to the overall receiver noise temperature, signal levels, total gain, noise power density, total noise power, and measures of dynamic range. This tool is helpful when performing system design and trade-off analysis. Figure 2 shows the end-to-end model developed for this RF system using this spreadsheet tool. The model includes every component of the system starting from the antenna all the way to the analog-to-digital converter. The group of components that constitutes the RID is shown shaded.

The RID frequency conversion scheme uses two frequency mixes. In the first mix, the entire 1- to 10-GHz block is upconverted using a high side LO (LO1). This LO is tuned over the range of 11.8 to 20.5 GHz with a maximum step size of 300 MHz. The IF is the lower sideband of the mix. The advantage of the high-side LO mix is that the LO harmonics generated by the mix, which create strong intermodulation products, are out of band. The upconversion is designed such that the IF band is outside the input RF band. Producing any signals in the receiver in the input band could lead to false-signal detection. The third constraint was to keep the frequencies as low as possible for reasons of cost and ease of implementation.

The second mix uses another high-side LO, thus restoring the inverted spectrum to its original orientation. The second LO (LO2) is fixed at a frequency of 11.2 GHz, which is simple to synthesize by phase-lock multiplying a 100-MHz pure reference signal by a factor of 112. The resulting output IF band is centered at 550 MHz with a bandwidth of 350 MHz. Since the IF bandwidth is narrower than an octave, it may be frequency converted as a block in the next stage of the receiver. Also, 550 MHz is a

practical IF frequency for transmission over distance using a coaxial cable or a fiber-optic link.

Although wideband devices that cover the wide input band of 1- to 10 GHz exist, the band needed to be narrowed before the first mix to meet the high degree of spectral purity required. A switched-filter-bank approach was selected with eight contiguous bands that divide the 1- to 10-GHz block optimally for minimizing mixer intermodulation products, physical size, and complexity. Figure 3 is a block diagram of the RID.

## IV. Local Oscillator Requirements

The requirements for the local oscillators of a down-converter are dependent on the specific frequency scheme and gain profile of the downconverter. For the RID to meet its phase noise, spectral purity, and dynamic range requirements, LO1 and LO2 must meet the requirements presented below. The requirement-derivation process is also described.

### A. Local Oscillator Power Level

The LO power level requirement results from the required mixer performance: the higher the LO power, the higher the mixer dynamic range and intermodulation suppression. It was determined that the LO1 level is required to be +10 dBm while the LO2 level is required to be +20 dBm. LO2 power is higher than that of LO1 because the gain stages between the two mixers result in increased signal levels and thus increased difficulty in meeting the dynamic range requirements.

### B. Spectral Purity

The spectral purity requirements result from analyzing all the different ways in which a spurious signal (spur) accompanying the LO could end up in the IF band and be falsely detected as a signal of interest. The required suppression of spurs relative to the LO is measured in dBc (dB relative to carrier). There are three frequency regions around the LO frequency that are derived separately [2]. For example, the requirements for LO2 are derived below. First, the frequency ranges of the second frequency conversion are listed:

- (1) LO2 frequency is 11,200 MHz.
- (2) Input RF range is 10,450 to 10,850 MHz.
- (3) Output IF range is 350 to 750 MHz.

The first spectrally pure region extends over an IF bandwidth (400 MHz) in each direction around LO2. Any spur frequency within this range will modulate an incoming signal in the same way it modulates the LO signal. The required suppression for this range is derived by subtracting the minimum detectable signal from the maximum signal of interest:

$$-73 - (-172) \text{ dBm} = 99 \text{ dB}$$

Therefore, for 10,800 to 11,600 MHz, the required suppression is  $-99 \text{ dBc}$ .

The second region includes any spur frequency that, when mixed with LO2, will end up in the IF band. After leaking from the L port to the R port in the mixer, the spur is required to be 3 dB below the weakest incoming signal of interest. Therefore, for 10,450 to 10,850 MHz and 11,550 to 11,950 MHz, the required suppression is  $-102 \text{ dBc}$ .

Finally, the third region consists of spur frequencies in the IF range that may leak from the L port to the I port in the mixer. After leaking from the L port to the I port, the spur is required to be 3 dB below the weakest signal of interest at the IF band. Therefore, for 350 to 750 MHz, the required suppression is  $-109 \text{ dBc}$ .

### C. Phase Noise

The receiver phase noise requirement (Table 1, no. 6) can be met only if the contributions from all the local oscillators do not add up to exceed it. Since no specific IF/BB design is discussed here, the LO contribution from the IF/BB downconverter will be omitted here.

LO1 and LO2 frequencies are derived from a common reference frequency using phase-locked multiplication techniques. The relative contribution from each LO depends on its frequency relative to some reference frequency. Within the phase-locked loop bandwidth, the phase noise of each LO is synchronously related to the reference phase noise by a factor of  $20 \times \log N$ , where  $N$  is the frequency multiplication factor [3]. The LO1 maximum frequency is 20,500 MHz, and the LO2 frequency is 11,200 MHz. Therefore, within the loop bandwidth, LO1 must exceed the receiver phase noise requirements by

$$20 \times \log((20,500 + 11,200)/20,500) = 3.8 \text{ dB}$$

LO2 must exceed the receiver phase noise requirements by

$$20 \times \log((20,500 + 11,200)/11,200) = 9.0 \text{ dB}$$

and a 100-MHz frequency reference has to exceed the receiver phase noise requirements by

$$20 \times \log(205 + 112) = 50 \text{ dB}$$

Outside the loop bandwidth, the phase noise of each LO is independently random. We may, for example, allocate equal contributions from each LO, such that LO1 and LO2 each has to exceed the receiver phase noise requirements by 3 dB.

## V. Implementation

### A. Circuit Description

We discussed the requirements and the design approach for the RID. The next step involves selecting circuit components and devices that implement the required performance called for by the design.

Figure 3 is a block diagram of the RID. A1 is a miniature wideband LNA, custom designed for high output power. A1 is followed by PA1, which is a programmable step attenuator covering the range of 0 to 11 dB in 1-dB steps. The attenuator can be used to compensate for a potentially long cable between the system LNA and the RID; this cable would cause insertion-loss variation with frequency. PA1 can also attain higher dynamic range (at a cost of increased system temperature) in the presence of very strong interfering signals. PA1 is computer monitored and controlled.

PA1 is followed by FB1, a computer-controlled, switched-filter bank implemented in microstrip with solid-state switches at the input and output. The solid-state switch approach has the advantages of higher reliability and smaller size as compared to the multiport coaxial electromechanical switches. These advantages are important in this system. The disadvantages of the solid-state switch as compared to the electromechanical switch are higher insertion loss, poorer impedance match, increased sensitivity to strong signals, and amplitude variation with temperature. All the disadvantages were adequately addressed. The switched-filter bank was designed to have adequate insertion loss, flatness, and immunity to strong signals. The impedance match was improved with attenuator pads, and amplitude drift can be minimized with temperature control.

FB1 is followed by a very wideband IF mixer (M1), which is followed by two identical bandpass filters (BPF1-A,B) with center frequencies ( $F_c$ ) of 10,650 MHz and bandwidths ( $BW$ ) of 400 MHz. The reason for filtering twice

is to achieve the high degree of rejection required to suppress leakage from LO1, which would mix with the LO2 frequency and result in spurs in the final IF band. The bandpass filter pair is followed by A2, an amplifier designed for a very flat frequency response. A2 is followed by a high power mixer (M2). The mixer is followed by a bandpass filter (BPF2) which establishes the IF output bandwidth. Additional gain is provided by A3, an amplifier designed for high dynamic range.

The LO1 and LO2 signals are each amplified in the RID using AL1 and AL2, respectively. This allows an input power level of 0 dBm, a requirement that can be easily achieved by most modern signal generators. LO2 is sampled through a 30-dB directional coupler (CPL2) with a diode detector (DET2). The detector is input to the monitor and control interface circuit (MCIC), which is a custom-designed printed circuit board. On the MCIC, the detector output is amplified and compared to a reference threshold, producing a digital indication of LO2 signal presence, which may be read by a computer. SW1 is an electromechanical coaxial switch that selects between the sampled outputs of LO1 and the IF output band. SW1 is computer controlled and monitored. The selected input is designed to be connected, directly outside the RID, to the power sensor of a computer-monitored power meter (PM). The reading of the LO1 power level can be used to adjust LO1 to the desired power level over its wide frequency band. The dc voltages used to power the active components in the RID are monitored in the MCIC and reported to the computer.

The directional coupler (CPLRS) at the output of the RID serves to inject a test signal for testing the fiber-optic link. Signal injection points for the RID are designed into the LNA package.

## B. Packaging

Once the detailed circuit design is complete, the mechanical design follows. The main goal in the RID mechanical design was to achieve a low-noise environment for the circuit. The chassis designed for the RID circuit is shown in Figs. 4 and 5. Several techniques were used to reduce signal leakage and crosstalk, which would compromise the receiver spectral purity.

Coaxial cables are used to connect devices that are individually packaged with coaxial connectors. This method reduces signal leakage in comparison with a microstrip layout.

To isolate inputs and outputs of devices in the signal path, each critical signal-conditioning function is confined

in a separate chamber. The chassis is machined from a single block of aluminum with chambers on the top and bottom sides. The signal is routed from chamber to chamber via coaxial feed-through adapters. The top and bottom covers are thick and flat, which facilitates sealing the box tightly by using closely spaced screws.

Isolators and pad attenuators are used to improve the impedance matching between filters, amplifiers, and mixer ports.

The dc voltages that enter the chassis are first filtered in their own chamber and then distributed to other chambers. To reduce signal leakage from one device to another via the dc voltage wires, feed-through capacitors are used to pass wires through the chassis walls. Each capacitor provides 60 dB of suppression to a 10-GHz signal. In addition, ferrite beads are used around each wire to further attenuate high-frequency leakage. Finally, the wire used has a special dielectric coating that provides additional high-frequency filtering.

## VI. Test Results

The gain profile model (Fig. 2) was verified by testing. Testing included frequency response, noise temperature, gain stages, 1-dB compression, mixing intermodulation products, and *s*-parameters. The noise temperature and gain profiles of the RID generated by the model are shown in Figs. 6 and 7, respectively. One way we evaluated the effectiveness of the implemented design techniques was to tune LO1 to the frequency of 11,750 MHz. If this frequency leaks through the receiver, it may mix with the LO2 frequency (11,200 MHz) and result in a spur at 550 MHz, which is the center of the output IF band. The spur was measured at a level of -106 dBm, which is 17 dB below the noise floor of the RF system at the same point. This successful rejection of the spur is a result of the following measures: double filtering between M1 and M2; improved impedance matching of the mixers and filters using isolators; dc line filtering of AL1, AL2, and A2 using feed-through capacitors, ferrite beads, and lossy dielectric wire; shielded coaxial cables connecting the individually shielded signal processing components; and sealed compartments to separate the different frequency functions.

## VII. Summary

A 1- to 10-GHz downconverter was designed, built, and tested. The requirements were listed, and the design process was described. Performance was verified in standalone bench testing using HP83712A synthesizers for LO1,

LO2, and RF input. The requirements described above in Section II were met. Only one channel was assembled and tested by the time the HRMS project was eliminated by Congress.

The RID can be useful for other RF-to-IF frequency-conversion applications featuring an input frequency range

of 1 to 10 GHz with an instantaneous bandwidth of 350 MHz and a dynamic range of 125 dB/Hz. The RID was designed as a part of a complex computer-controlled RF system; however, it may be used as a stand-alone downconverter using commercial synthesizers for LO1 and LO2. A user interface box was constructed, facilitating local manual monitor and control functions.

## References

- [1] G. A. Zimmerman, M. F. Garyantes, M. J. Grimm, and B. Charny, "A 640-MHz 32-Megachannel Real-Time Polyphase-FFT Spectrum Analyzer," *The Telecommunications and Data Acquisition Progress Report 42-107, vol. July-September 1991*, Jet Propulsion Laboratory, Pasadena, California, pp. 132-140, November 15, 1991.
- [2] W. F. Egan, *Frequency Synthesis by Phase Lock*, Chapter 4, Malabar, Florida: Robert E. Krieger Publishing Co., pp. 61-75, 1990.
- [3] V. Manassewitsch, *Frequency Synthesizers Theory and Design*, New York: John Wiley & Sons, pp. 120-124, 1987.

**Table 1. RID requirements.**

No.	Parameter	Value
1	Input frequency range, GHz	1 to 10
2	Output instantaneous 0.5-dB bandwidth, MHz	320
3	Noise-temperature contribution of downconverter to front end LNA	5 percent max (1.2 K) for nominal attenuator settings
4	Two simultaneous channels with equal tuning frequency	
5	Computer controlled and monitored	
6	Receiver single-sideband phase noise	
	Frequency offset from carrier, Hz	Single-sideband phase noise in 1-Hz BW, dBc
	5	-49
	10	-64
	60	-97
	100	-105
	120	-105
	180	-107
	$\geq 300$	-112
7	Spectrally pure dynamic range for a single tone (of -73 dBm at antenna)	89 dBc at 10 GHz and 99 dBc at 1 GHz (dB from the tone down to mixing products with local oscillators and their spurs)
8	Nominal dynamic range (defined from noise floor to 1-dB compression point)	40 dB, for 320-MHz bandwidth; for a high dynamic range mode, 50-dB dynamic range is required and system noise temperature ( $T_{sys}$ ) is allowed to degrade by 3 K
9	Minimum gain, dB	34
10	Programmable gain range, dB	10; step size: 1 dB
11	Gain deviation (instantaneous peak to peak over 320 MHz), dB	$\pm 2$

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(a)
program IMODP.EXE by Dalia McWatters, output
file: INTERMOD.FRQ
LISTING OF INTERMODULATION PRODUCTS SUPPRESSION
RESULTING FROM THE FOLLOWING FREQUENCY CONVERSION:

FRL = 1000.000 FRH = 1600.000
RF FREQ. INCREMENT = 50.000
FLL = 11650.000 FLH = 12250.000
LO FREQ. INCREMENT = 50.000
IF = 10650.000
IF FILTER SHAPE = 40.0 dB @ 1.200 3dB bw from center freq.
BW = 400.000
REQUIRED REJECTION: 60.0dB
HIGHEST ORDER INTERMOD CALCULATED: 9
MIXER INTERMOD TABLE USED: LRL.L.TBL

F(M,N) = ABS(M*FR + N*FL) if in BW
-----

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(b)
program IMODP.EXE by Dalia McWatters, output
file: INTERMOD.FRQ
LISTING OF INTERMODULATION PRODUCTS SUPPRESSION
RESULTING FROM THE FOLLOWING FREQUENCY CONVERSION:

FRL = 1000.000 FRH = 1600.000
RF FREQ. INCREMENT = 100.000
FLL = 11650.000 FLH = 12250.000
LO FREQ. INCREMENT = 100.000
IF = 10650.000
IF FILTER SHAPE = 40.0 dB @ 1.200 3dB bw from center freq.
BW = 400.000
REQUIRED REJECTION: 99.0dB
HIGHEST ORDER INTERMOD CALCULATED: 9
MIXER INTERMOD TABLE USED: LRL.L.TBL

F(M,N) = ABS(M*FR + N*FL) if in BW
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FLO = 11650.000 FRF = 1000.000
Atten = 61.79dB
F( OR , -1L ) = 11650.000

FLO = 11650.000 FRF = 1000.000
Atten = 61.79dB
F( OR , 1L ) = 11650.000

FLO = 11650.000 FRF = 1100.000
Atten = 61.79dB
F( OR , -1L ) = 11650.000

FLO = 11650.000 FRF = 1100.000
Atten = 61.79dB
F( OR , 1L ) = 11650.000

FLO = 11650.000 FRF = 1200.000
Atten = 90.00dB
F( -9R , 0L ) = 10800.000

FLO = 11650.000 FRF = 1200.000
Atten = 61.79dB
F( OR , -1L ) = 11650.000

FLO = 11650.000 FRF = 1200.000
Atten = 61.79dB
F( OR , 1L ) = 11650.000

FLO = 11650.000 FRF = 1200.000
Atten = 90.00dB
F( 9R , 0L ) = 10800.000

FLO = 11650.000 FRF = 1300.000
Atten = 97.14dB

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Fig. 1. Mixing products analysis program outputs: (a) 60-dB suppression required and (b) 99-dB suppression required.

DOWNCONVERTER GAIN AND NOISE TEMP PROFILE											
with PROTOTYPE IF/BB				Dalla McWatters			12/28/93				
HRMS RF/IF				HRDC.XLS							
Noise Temp Kelvin	GAIN dB	Pol dBm	BANDWIDTH Hz	sigout dBm	Tsys Kelvins	TOTAL GAIN(dB)	PnOUT/Hz dBm/Hz	Pn(out) dBm	DR (dB)	HR (dB)	BLOCK
*****				*****				*****			
ANTENNA	14.4	0	7	2.50E+09	-62	14.4000	0.0	-187.0	-93.1		ANTENNA
PAD		0	7	2.50E+09	-62	14.4000	0.0	-187.0	-93.1		PAD
LNA	9.6	52	7	2.50E+09	-10	24.00	52.0	-132.8	-38.8	45.8	17.0 LNA
CABLE+PAD		-5	7	2.50E+09	-15	24.00	47.0	-137.8	-43.8		CABLE+PAD
A1	290	28	21	2.50E+09	13	24.01	75.0	-109.8	-15.8	36.8	8.0 A1
PA1		-5	21	2.50E+09	8	24.01	70.0	-114.8	-20.8		PA1
amp		0	21	2.50E+09	8	24.01	70.0	-114.8	-20.8		amp
FBI1+PADS		-16	21	5.00E+08	-8	24.01	54.0	-130.8	-43.8		FBI1+PADS
M1	1164	-7	-3	5.00E+08	-15	24.02	47.0	-137.8	-50.8	47.8	12.0 M1
BPF1		-3	-3	5.00E+08	-18	24.02	44.0	-140.8	-53.8		BPF1
A2	440	24.5	24	5.00E+08	6.5	24.04	68.5	-116.3	-29.3	53.3	17.5 A2
PAD		-0.5	24	5.00E+08	6	24.04	68.0	-116.8	-29.8		PAD
M2	2014	-9	8	5.00E+08	-3	24.04	59.0	-125.8	-38.8	46.8	11.0 M2
BPF2+PADS		-12	8	5.00E+08	-15	24.04	47.0	-137.8	-50.8		BPF2+PADS
A3	360	40	30	5.00E+08	25	24.05	87.0	-97.8	-10.8	40.8	5.0 A3
CPLRS+PAD		-7	30	5.00E+08	18	24.05	80.0	-104.8	-17.8		CPLRS+PAD
FOL	6000000	-17	1	5.00E+08	1	24.11	63.0	-121.8	-34.8	35.8	0.0 FOL
pad		-15		5.00E+08	-14	24.1162	48.0	-136.8	-49.8		pad
IF/BB D/C in		0		1.10E+08	-14	24.1162	48.0	-136.8	-56.4		IF/BB D/C in
LPF + pads		-4		1.10E+08	-18	24.1231	44.0	-140.8	-60.4		LPF + pads
Mixer1	1200.0	-6	7	1.10E+08	-24	24.1709	38.0	-146.8	-66.4	73.4	31.0 Mixer1
BPF + circ		-2		1.10E+08	-26	24.1978	36.0	-148.8	-68.4		BPF + circ
AMP2	130.0	23	5	1.10E+08	-3	24.2304	59.0	-125.8	-45.4	50.4	8.0 AMP2
Mixer2 + pad		-19	-3	1.10E+08	-22	24.2591	40.0	-144.8	-64.4	61.4	19.0 Mixer2 + pad
AMP3	130.0	37	15	1.10E+08	15	24.2721	77.0	-107.8	-27.4	42.4	0.0 AMP3
SAW + pads		-42		4.00E+07	-27	24.3638	35.0	-149.8	-73.7		SAW + pads
AMP4	130.0	37	15	4.00E+07	10	24.4049	72.0	-112.7	-36.7	51.7	5.0 AMP4
MIXER3 + pad		-14	12	4.00E+07	-4	24.4053	58.0	-126.7	-50.7	62.7	16.0 MIXER3 + pad
AMP5	7000.0	20	29	4.00E+07	16	24.4164	78.0	-106.7	-30.7	59.7	13.0 AMP5
lpf + pads + PA2	(-26 to -57)	-32		4.00E+07	-16	24.4237	46.0	-138.7	-62.7		lpf + pads + PA2
AMP6	7000.0	20	29	4.00E+07	4	24.5995	66.0	-118.7	-42.7	71.7	25.0 AMP6
PAD		-2		4.00E+07	2	24.5996	64.0	-120.7	-44.7		PAD
AMP7	7000.0	21	29	4.00E+07	23	24.6024	85.0	-99.7	-23.7	52.7	6.0 AMP7
PAD		-15		4.00E+07	8	24.6024	70.0	-114.7	-38.7		PAD
AMP8	7000.0	21	29	4.00E+07	29	24.6031	91.0	-93.7	-17.7	46.7	0.0 AMP8
PAD		-3		4.00E+07	26	24.6031	88.0	-96.7	-20.7		PAD
IF/BB D/C out		0		4.00E+07	26	24.6031	88.0	-96.7	-20.7		IF/BB D/C out
WBSA PA		-28		4.00E+07	-2	24.6034	60.0	-124.7	-48.7		WBSA PA
CLC100	865.0	20	12	4.00E+07	18	24.6042	80.0	-104.7	-28.7	40.7	-6.0 CLC100
pad		-3		4.00E+07	15	24.6042	77.0	-107.7	-31.7		pad
ADC INPUT	14500000.0	0	-1	4.00E+07	15	24.8936	77.0	-107.7	-31.6	30.6	-16.0 ADC INPUT

**Fig. 2. Receiver gain and noise profile spreadsheet.**



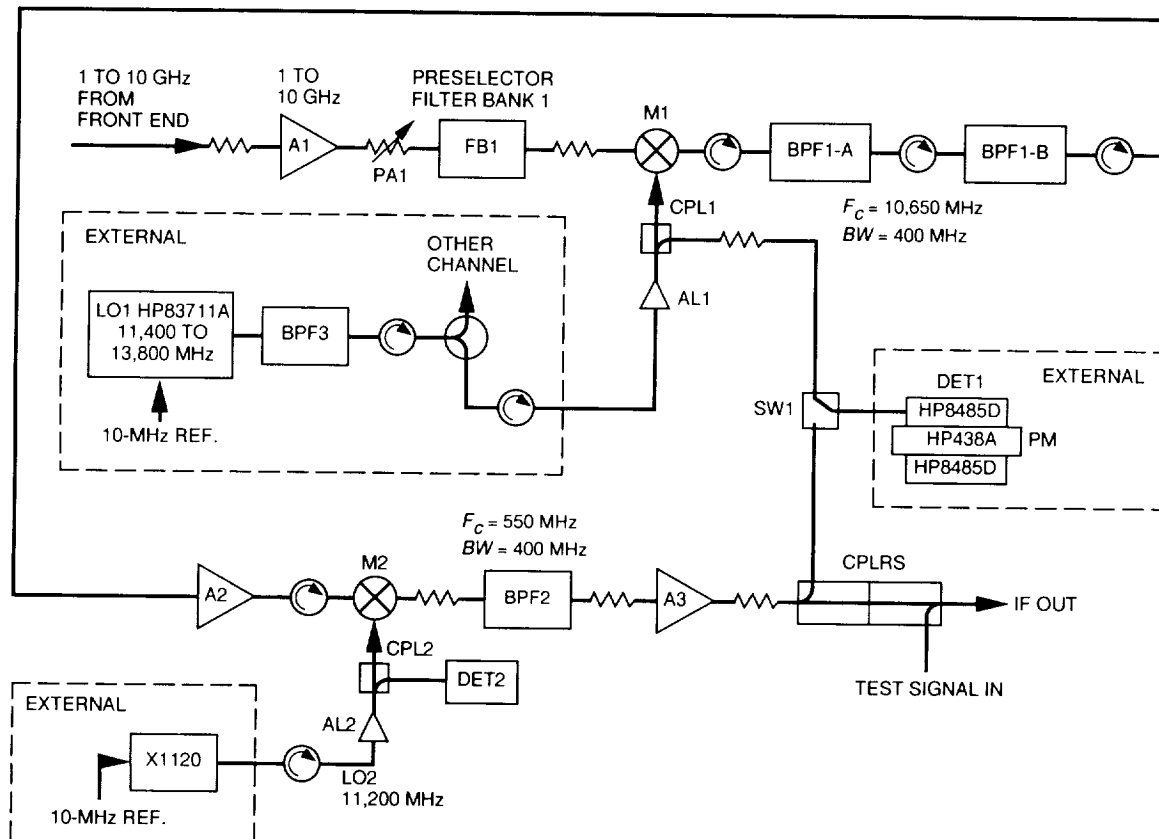


Fig. 3. RF/IF downconverter block diagram.

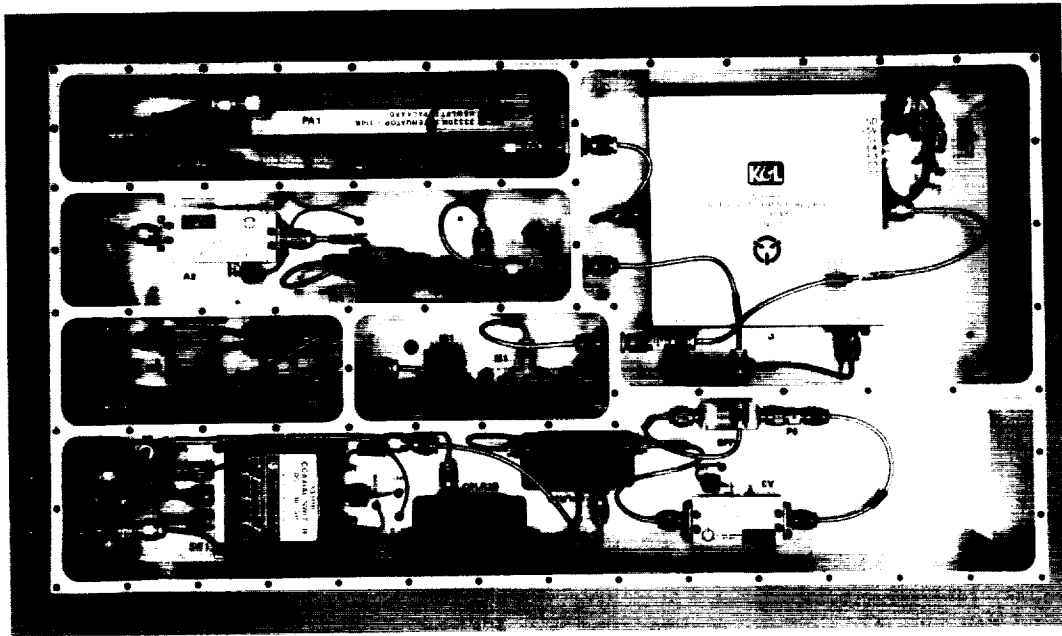


Fig. 4. RF/IF downconverter chassis fully assembled, top side.

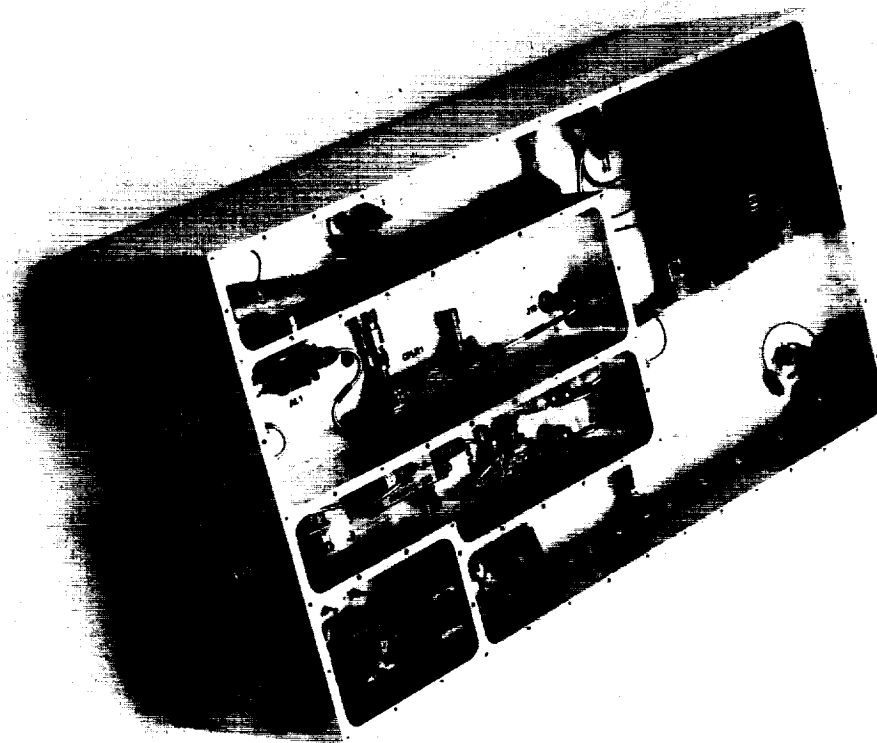


Fig. 5. RF/IF downconverter chassis fully assembled, bottom side.

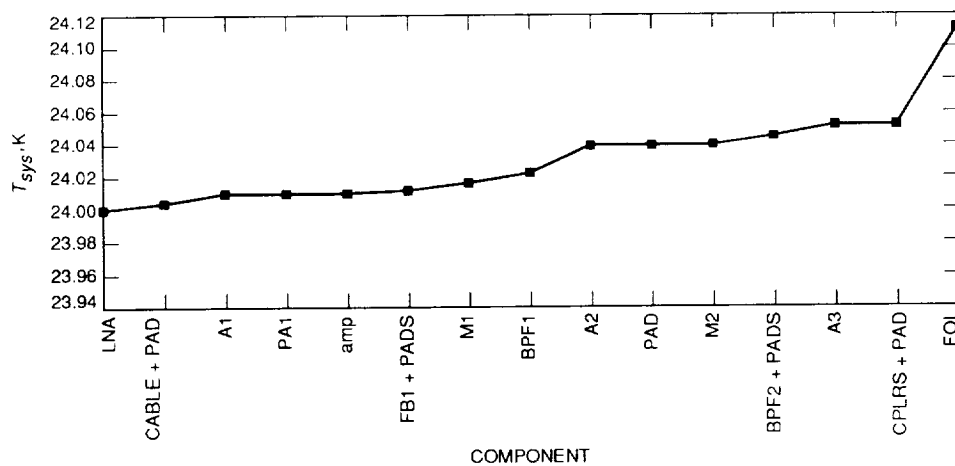


Fig. 6. RF/IF downconverter cumulative system temperature by component.

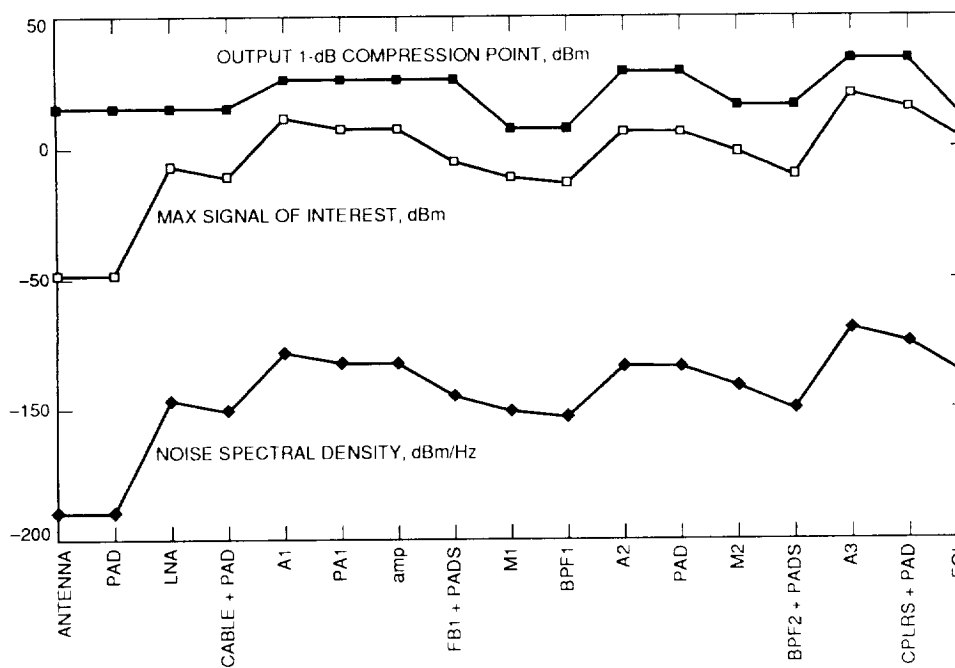


Fig. 7. RF/IF downconverter gain profile.

# A Closed-Loop Time-Alignment System for Baseband Combining

Y. FERIA

Communications Systems Research Section

*In baseband combining, the key element is the time alignment of the baseband signals. This article describes a closed-loop time-alignment system that estimates and adjusts the relative delay between two baseband signals received from two different antennas for the signals to be coherently combined. This system automatically determines which signal is advanced and delays it accordingly with a resolution of a sample period. The performance of the loop is analyzed, and the analysis is verified through simulation. The variance of the delay estimates and the signal-to-noise ratio degradation in the simulations agree with the theoretical calculations.*

## I. Introduction

Antenna array combining techniques have been used in the Deep Space Network to improve the signal-to-noise ratio (SNR) [1]. These techniques include full spectrum, baseband, and symbol stream combining. Each technique, however, has been applied only to a specific mission, and a rigorous comparison between the techniques is missing. To have a better understanding of how these techniques perform under the same or different conditions, all techniques are being studied and simulated in software.

This article concentrates on baseband combining where signals from two antennas have each had their carrier removed before going into the combiner. A diagram of baseband (BB) combining is shown in Fig. 1. The signals received from the  $i$ th antenna have the following form [1]:

$$\begin{aligned} r_i(t) = & \sqrt{2P_{C_i}} \sin(\omega_{c_i}(t - \tau_i) + \theta_{c_i}) + \sqrt{2P_{D_i}} d(t - \tau_i) \\ & \times \operatorname{sgn}[\sin(\omega_{s_c}(t - \tau_i) + \theta_{s_c})] \\ & \times \cos(\omega_{c_i}(t - \tau_i) + \theta_{c_i}) + n_i(t) \end{aligned} \quad (1)$$

where  $P_{C_i} = P_i \cos^2 \Delta$  and  $P_{D_i} = P_i \sin^2 \Delta$ , with  $P_i$  being the total signal power of the received signal from the  $i$ th antenna, and  $\Delta$  is the modulation index. The parameters  $\omega_{c_i}$  and  $\theta_{c_i}$  denote the carrier angular frequency and phase of the received signal from the  $i$ th antenna,  $\omega_{s_c}$  and  $\theta_{s_c}$  denote the subcarrier angular frequency and phase, and  $\tau_i$  is the relative delay between the reference signal and the  $i$ th signal. These received signals are analog-to-digital (A/D) converted and downconverted to an intermediate frequency (IF) and then sampled before going through the